Optically Transparent Ground Planes with High Order Response Using Miniaturized Element Frequency Selective Surfaces

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Abstract

In this paper design of a multi-pole wide band optically transparent ground plane is considered. Challenges of the design by stacking multiple layers of single pole miniaturized element frequency selective surfaces are explained. An alternate arrangement of the layers and spacing between the layers is proposed. A three-pole transparent ground plane is designed band fabricated based on the proposed technique. The performance of the designed ground plane using analytical and computational methods is evaluated experimentally and excellent agreement is demonstrated.

1 Introduction

An optically transparent ground plane is needed for many applications such as transparent antenna systems or electromagnetically shielded windows. It was recently shown that miniaturized element frequency selective surfaces (MEFSS) can be utilized to design efficient optically transparent ground planes[1]. MEFSSs are typically planar structures consisting of an array of sub-wavelength metallic unit cells. The main advantage of MEFSSs over the conventional frequency selective surfaces is that MEFSSs have a much smaller unit cell size compared to the wavelength which allows full analysis of the structure using a lumped element circuit model[2]. The circuit model of a single pole MEFSS with a stop-band is that of a one stage band-stop filter with shunt components. According to standard filter theory, the order of the filter can be increased by adding series and shunt stages alternatively. However, fabrication of series components between two layers of MEFSS is extremely difficult. A commonly used alternative method of realizing series components is to use a $\lambda/4$ long spacer as an impedance invertor between the two layers which can be modeled as a transmission line with the same length. The transmission line will effectively transfer the shunt elements into series elements as far as the input impedance of the filter is concerned [3]. However, this approximation is only accurate for narrow band filters (10% or lower) since the electrical length of the transmission line varies with frequency. Moreover, the thickness of the MEFSS increases linearly by increasing the number of poles which in turn increases the angle dependency of the phase shift between consecutive layers and therefore the angle dependency of the response of the MEFSS. Another difficulty in implementing this method for MEFSSs is that the value of shunt elements needed will change dramatically from one layer to the next which is often not feasible due to fabrication difficulty or the fact that the unit cell size must remain small compared to the wavelength. In the next section an alternative method of designing a wide-band multi-pole MEFSS is discussed.

2 FSS analysis and design

The circuit model and a unit cell of a single pole MEFSS consisting of an array of metallic loops is shown in Fig.1. The Capacitor models the gap between the adjacent sides of the metallic squares that are perpendicular to the direction of the incoming electric field at a given polarization and the inductor models the sides of the metallic square that are perpendicular to the direction of the incoming magnetic field. The transmission lines model the propagation of electromagnetic waves in the free space and the dielectric substrate on which metallic loops are placed.

A multi-pole response can be achieved by stacking multiple layers of single pole MEFSSs. The mentioned challenges in designing a wide-band multi-pole FSS can be avoided by allowing different layers of the cascade

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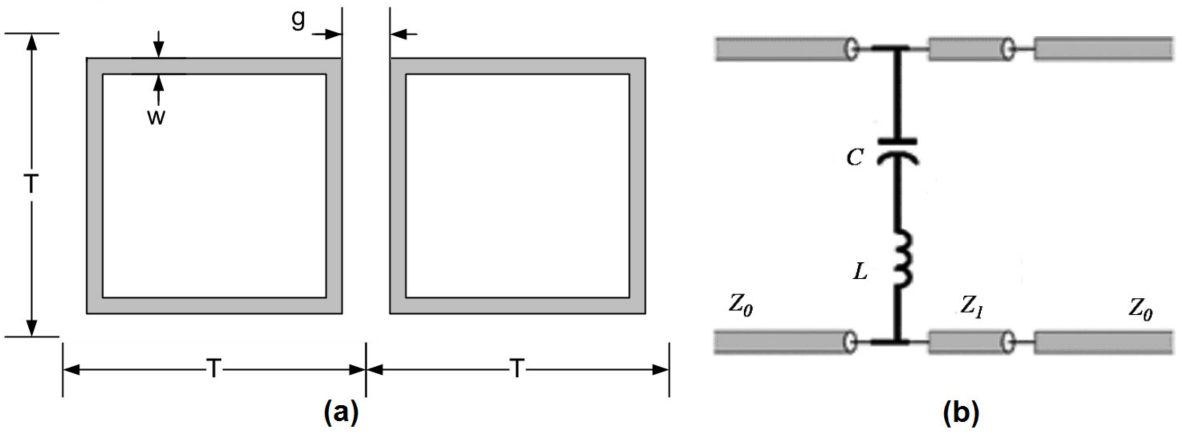

Figure 1: (a) Two adjacent unit cells of the FSS. (b) The equivalent circuit model of the FSS.

FSS to have different resonant frequencies. Each FSS layer can be modeled with a shunt admittance Y_n. If the layers are stacked very close together, the entire N-layer structure can be modeled with a single shunt admittance $Y = \Sigma_{n=1}^{N} Y_n$. The normalized shunt admittance can be written in the form of the ratio of two polynomials of ω^2:

$$\bar{Y} = \frac{Y}{Y_0} = \frac{j\omega P(\omega^2)}{Q(\omega^2)} \tag{1}$$

The reflection coefficient of the FSS follows:

$$\Gamma = \frac{1-(1+\bar{Y})}{1+(1+\bar{Y})} = \frac{-j\omega P(\omega^2)}{2Q(\omega^2)+j\omega P(\omega^2)} \tag{2}$$

The normalized admittance of each layer has the form,

$$\bar{Y_n} = \frac{j\omega P_n(\omega^2)}{Q_n(\omega^2)} = \frac{k_n j\omega}{\omega_n^2 - \omega^2} \tag{3}$$

where $k_n = Z_0/L_n$ and $\omega_n^2 = 1/L_n C_n$. Therefore,

$$\begin{aligned} Q(\omega^2) &= \prod_{n=1}^{N} (\omega_n^2 - \omega^2) \\ P(\omega^2) &= \sum_{m=1}^{N} k_m \prod_{n \neq m} (\omega_n^2 - \omega^2) \end{aligned} \tag{4}$$

And the magnitude of the reflection coefficient can be written as:

$$|\Gamma|^2 = \frac{\omega^2 |P(\omega^2)|^2}{4|Q(\omega^2)|^2 + \omega^2 |P(\omega^2)|^2} \tag{5}$$

It is evident that $|\Gamma| = 1$ when $Q(\omega^2) = 0$ and $|\Gamma| = 0$ when $P(\omega^2) = 0$. $Q(\omega^2)$ has exactly N zeros at $\omega^2 = \omega_n^2$. $P(\omega^2)$ can be expressed as

$$P(\omega_n^2) = k_n \prod_{m=1}^{n-1} (\omega_m^2 - \omega_n^2) \prod_{m=n+1}^{N} (\omega_m^2 - \omega_n^2) \tag{6}$$

Assuming ω_n^2 are indexed in ascending order, all the terms in the first product in (6) are negative and all the terms in the second product are positive which yields:

$$\begin{aligned} P(\omega_n^2) > 0 \quad &\text{for even n} \\ P(\omega_n^2) < 0 \quad &\text{for odd n} \end{aligned} \tag{7}$$

Therefore the sign of $P(\omega^2)$ changes between any two consecutive zeros of $Q(\omega^2)$. Hence $P(\omega^2)$ has at least one zero between ω_n^2 and ω_{n+1}^2. Moreover, since $P(\omega^2)$ is of order $N-1$, it must have exactly one zero between ω_n^2 and ω_{n+1}^2. Therefore the response of the filter is consisted of N stop-bands and N-1 pass-bands. There are $2N$ unknowns, C_n and L_n, that have to be determined to complete the filter design. Poles, ω_n^2, and zeros, ν_m^2, of the entire filter can be selected independently except that there should be one zero between each two consecutive poles. The values of k_n, ω_n^2 and ν_m^2 have the following relationship:

$$\mathbf{A}k = b \tag{8}$$

where

$$\begin{aligned}
[k]_m &= k_m \\
[b]_m &= -k_N \prod_{n=1}^{N-1} (\omega_n^2 - \nu_m^2) \qquad\qquad m = 1, 2, ..., N-1 \\
[\mathbf{A}]_{m,n} &= \prod_{i \neq n}^{N} (\omega_i^2 - \nu_m^2) \qquad\qquad\qquad n = 1, 2, ..., N-1
\end{aligned}$$

Once ω_n^2 and ν_m^2 are selected, k_1 to k_{N-1} can be found as a factor of the free parameter k_N using equation (8). The existence of $N-1$ pass bands in the response is clearly unwanted. The pass bands can be neglected by separating the layers by a dielectric spacer. In practice one such spacer is sufficient to eliminate all the pass bands provided that a certain arrangement of the layers is used. Consider two subsets of layers with responses $Y_1 = j\omega R(\omega^2)/S(\omega^2)$ and $Y_2 = j\omega T(\omega^2)/U(\omega^2)$. The subsets are selected such that zeros of S, s_i^2 and zeros of T, t_i^2 are equal to ω_{2i-1}^2 and ω_{2i}^2 respectively. Based on the above discussion, zeros of R, r_i^2 can be selected arbitrarily as long as the following inequality holds: $s_i^2 < r_i^2 < s_{i+1}^2$. Since U has a zero between each two consecutive zeros of S, it is possible to select k_{2i} such that zeros of R are also zeros of U. Similarly, k_{2i-1} can be selected such that zeros of T are also zeros of S. with Γ_2 denoting the reflection coefficient in presence of only Y_2, the response of the entire filter can be written as

$$|\Gamma|^2 = \left| \frac{2\Gamma_2 R - (1+\Gamma_2)R}{2S + (1+\Gamma_2)S} \right|^2 = \frac{\omega^2 |RU - ST|^2}{|2SU|^2 + \omega^2 |RU + ST|^2} \tag{9}$$

This result has to be consistent with that of equation (5). Therefore zeros of $(RU - ST)$ must be equal to ν_n. If the two subsets of layers are separated using a spacer such that the reflection coefficient due to the second subset can be written as $\Gamma_2' = \Gamma_2 exp(-j2\theta)$, the transmission in pass-bands will be suppressed. Assuming a constant θ of $\pi/2$, the magnetite of reflection coefficient can be written as

$$|\Gamma|^2 = \frac{\omega^2 |RU - ST|^2 + \omega^4 |RT|^2}{|2SU + \omega^2 RT|^2 + \omega^2 |RU + ST|^2} \tag{10}$$

The reflection coefficient vanishes at a certain frequency only if both $(RQ - ST)$ and (RT) are zero at that frequency. This is not possible since zeros of $(RQ - ST)$ are ν_n^2 and zeros of (RT) are selected to be ω_n^2. In practice the phase shift between the two layers depends on the frequency. However a minimum spacer thickness can often be found so that the transmission through the FSS stays below a desired value over the entire bandwidth.

The above mentioned technique was used to design and fabricated a three layer transparent ground plane with insertion loss of more than 10dB over the frequency range of 3GHz to 8.9Ghz. Each layer consists of an array of square loops as shown in Fig.1. All three layers share the same line thickness $w = 0.2mm$, and periodicity of $T = 8mm$ which transfers to optical opacity of almost 30%. The gap sizes of the three layers are $g_1 = 1mm$,$g_2 = 0.25mm$ and $g_3 = 0.5mm$. The spacing between the first and second layer is filled with a polyester sheet with the thickness of $d_1 = 0.25mm$ and dielectric constant of 3.2. A slab of glass with the thickness of $d_2 = 5mm$ and dielectric constant of 5.7 is used between the second and third layers. An optimization is performed to minimize d_2 while maintaining a return loss of more than $20dB$ in the frequency range between the first and last resonance. The value of d_2 is close to $\lambda/4$ at the center of the stop-band.

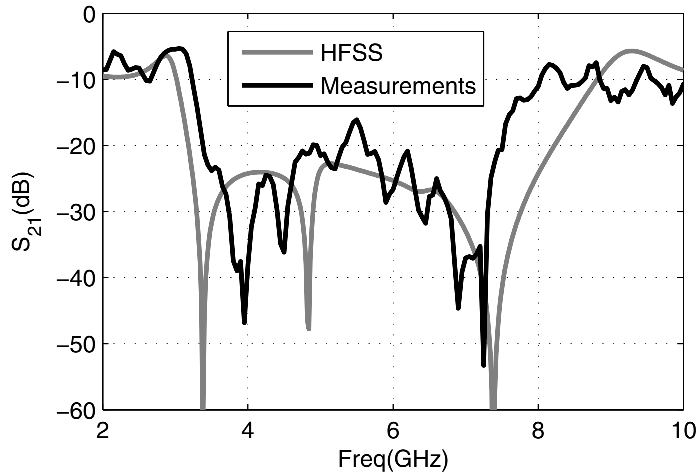

Figure 2: Measurement resutls

A 30cm by 30cm sheet with the above specifications was fabricated and its insertion loss was measured was measured by placing the antenna in the line of sight of two horn antennas and measuring the drop in the transmission between the two antennas. Fig.2 shows the measured and simulated transmission coefficient of the surface. The slight difference between the measurement and simulation results is due to the trapezoidal shape of the cross section of the metallic traces which reduces the capacitance between the adjacent metallic traces.

3 Conclusion

A novel technique in designing a multi-pole stop-band miniaturized element frequency selective surfaces using lower order MEFSSs was introduced. It was shown that the proposed method allows achieving a wide-band multi-pole response while maintaining a relatively small thickness, allowing to reduce the angle dependency of the multi-pole MEFSS. A three pole transparent ground plane was designed using the proposed method and a 10dB insertion loss bandwidth of almost 3 to 1 was achieved.

4 References

1. M. Kashanianfard and K. Sarabandi, “A transparent ground plane using miniaturized element frequency selective surfaces,” *2010 IEEE International Symposium on Antennas and Propagation*, 11-17 July 2010

2. K. Sarabandi and N. Behdad, “A frequency selective surface with miniaturized elements,” *IEEE Trans. Antennas Propag.*, vol. 55, pp. 12391245, May 2007.

3. D.M. Pozar. *Microwave Engineering.* Amherst, MA: John Wiley & Sons, Inc., 2005, pp. 411-431